BROWN UNIVERSITY Providence, Rhode Island • 02912

DIVISION OF ENGINEERING

April 20, 1979



NASA Scientific and Technical Information Facility
P. O. Box 8757
Baltimore/Washington International Airport
Maryland 21240

Re: Final Contract Report
for the period 1 August 1975 to 30 April 1977
NASA Grant No. NSG 1222
"Methods of Fabrication of Large Photodiode/Amplifier Arrays"
William R. Patterson, III, Principal Investigator

Gentlemen:

The subject contract was conceived, proposed and funded as part of a larger program under the sponsorship of the Langley Research Center Flight Instrument Division Section headed by Mr. F. O. Huck to develop new hardware technology and new engineering criteria for mechanical scanning spacecraft cameras. Cameras of this type have been used on several missions from the Soviet Union and, most notably, by the two Viking Landers which the United States placed on Mars in the summer of 1976. In references one and two we have discussed the design and performance of the Viking cameras. It was felt that should there be a follow on mission to Viking in the 1980's, the engineering rationale for the use of this type of camera might still be compelling if the wavelength range of the cameras could be extended into both the near ultraviolet (280-400 nm) and the near infrared (1000-2500 nm). In addition, the experiences of Viking suggested a need for simpler methods to fabricate the photosensor elements in such cameras, a need for new optical train components, and a need to systematically study certain other design options.

The program established to address these needs included this grant at Brown University, a contract with the Microelectronics Laboratory of Martin Marietta Aerospace Corporation, Denver, Colorado for an infrared sensor array and a contract with the Itek Corporation, Lexington, Massachusetts for new lens design. The Brown University grant contemplated two basic tasks. The first was to develop specialized silicon photodiodes optimized for use in mechanical scanning cameras through careful apertture design, integral optical masking, and enhancement of ultraviolet sensitivity. The second was to develop a special purpose, integrated circuit chip in the form of a power switchable, wide band, FET input, operational amplifier.

(NASA-CR-158525) METHODS OF FABRICATION OF LARGE PHOTODIODE/AMPLIFIER ARRAYS Final Report, 1 Aug. 1975 - 30 Apr. 1977 (Brown Univ.) 30 p

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Very early in this grant it was mutually agreed by the sponsoring group and the principal investigator that the second task was premature in terms of the overall program. Instead, several substitute tasks were decided upon. In addition to photodiode development, the final scope of work included direct participation in the design of the infrared sensor to be built by Martin Marietta, and preparation of two engineering studies directed at optimization of the Modulation Transfer Function and of the digitization levels in such cameras. We outline below the work actually accomplished, leaving some specific results to the appendices.

The contract with Martin Marietta and this grant began at the same time and ran for approximately the same period. Appendix I contains position papers prepared by the principal investigator under this grant as to the appropriate design of an infrared photosensor array based on lead sulfide detectors. Appendix I-A is a letter to Mr. V. F. Young outlining a fairly complete concept of such an array. Appendix I-B is a supplement to that letter and discusses the trade-off in signal to noise and drift involved in the use of an integral high speed (800 Hz) optical chopper in such an array. The signal to noise study resulted from continuing discussions between the principal investigator, Mr. Young and workers at Langley Research Center. The study showed that while a chopper has definite advantages, it is not as overwhelmingly effective as we at first thought it would be. Throughout the grant the principal investigator continued to participate in design reviews of the MMA work, which was strongly influenced by his participation.

Appendix II presents results that were derived as part of efforts to optimize design criteria for mechanical scanning cameras. It discusses the relation of detector noise and the quantization noise of the system analog to digital converter. In particular it addresses the extent to which noise can be reduced by averaging and how the digital signal is to be interpreted. The treatment is significantly different from that found in standard textbooks (see, for example, ref. 3) and does not seem to be available in the open literature.

Studies of the relationship between the spatial power spectra and auto-correlation function of images acquired by mechanical scanning cameras and those of conventional film cameras were also done. Given the power spectrum of an image which is approximately circularly symmetric as measured by one type of camera, results of this work enable one to calculate easily the spectrum to be expected from the other type of camera. The results are not entirely original (for example, ref. 4) but they do not seem well known. They were derived as the first step in making the criteria for optimum selection of MTF and sampling rate more scientifically based than they were during design of the Viking cameras. This part of the work is continuing under other grants.

The goal for photodiode construction was to build three types of diodes all having a common metalization pattern. After discussion with the contract monitor, the aperture pattern shown in Fig. 1 was agree on. The diode size and separation fit the scanning camera belonging to the Flight Instrument Division. The three aperture patterns provide a simple method of comparing results from different shapes. Surrounding each aperture is a region of black chrome plating to reduce specular reflection from the gold top metalization. This is a technique suggested to eliminate the calibration problems encountered in the Viking cameras which were caused by internal reflections in the photosensor array. That problem is discussed in ref. 1.

The first type of diode to be constructed is a simple p-boron diffusion in a n-substrate similar to the Viking photodiodes. These diodes would allow direct comparison of the improved aperture shapes with the Viking diode. The Viking photodiode proved to be sensitive to nuclear radiation from the radioisotope power sources on the lander. Attempts in that program to reduce sensitivity by using n-phosphorus diffusion in a p-substrate failed due to excessive surface leakage currents. The second photodiode type to be built under this grant is an n-diffused diode in a p-substrate but having a $p^+$-diffused guard band around the diode and having the masking oxide grown in a chlorine atmosphere. Figure 2 shows the proposed structure. These two measures should reduce the surface leakage current to negligible levels.

The third type of photodiode is directed at the problem of small area, high performance diodes in the near ultraviolet wavelength range. Devices for these wavelengths must have high internal quantum efficiency and low surface reflectivity. It is difficult to make diffused devices to have the high internal efficiency. Also diffused devices then need an extra antireflection coating because the reflectivity of bare silicon is very high (about 60%) in the ultraviolet. We propose an MOS device which makes use of the natural charge of an oxide grown at relatively low temperatures to invert the surface of 100 ohm·cm, p-type silicon. Figure 3 shows the proposed structure. Judicious choice of oxide thickness provides a built-in antireflection coating. The technique of MOS diodes without thickness optimization has been used for large area devices (ref. 5). We propose a $p^+$ guardband diffusion to limit the sensitive areas.

The following steps were completed in the grant period towards the goal of building these devices:

1. Black chrome plating on wafers is a new process for this laboratory and a small scale, plating system with regulated, low current sources and with temperature regulation was built.

2. Platinum-titanium-gold contacts are also a new technique for us. Materials were purchased and experiments performed to standardize procedure for making such contacts.

3. A suitable set of photo masks for the microlithography were designed. These were then cut two hundred times oversize in Rubylith, reduced photographically and a final mask set with an 11 x 11 pattern of devices was made.

4. Chlorine doping of oxides is a new process in this laboratory. We began but did not complete under this grant the development of standard procedures for this.

5. The necessary silicon wafers and processing chemicals were purchased and a preliminary run of diodes of the first type was done. However, no functional devices resulted because of a mask alignment error.

Sincerely yours,

Carl Cometta

Carl Cometta, Executive Officer
Division of Engineering

William R Patterson III

William R. Patterson, III
Senior Research Engineer

WRP/jsl
Enclosures

Appendix I-A

BROWN UNIVERSITY *Providence, Rhode Island • 02912*

DIVISION OF ENGINEERING

January 23, 1976

Mr. Virgil F. Young
Martin Marietta Aerospace
P. O. Box 179
Denver, Colorado

Dear Virg:

This will serve to make more clear and concrete some of the comments I made to you last month regarding the proper design criteria for the infrared detector assembly your group is planning to assemble for the Flight Instrument Division of LRC. As I said then, I am concerned that each of our projects should have a definite focus on particular aspects of advancing the technology of scanning camera detector arrays. That, after all, is what the bucks are for. It is also important that the planning process not take so long that it erodes the time and dollars available for actual hardware. To this end I am sending copies of this letter to Messrs Huck and Kelly at Langley, and I suggest that in a week or so when everyone has had a chance to digest it, we have a conference call to decide which of my suggestions are germane and which are not. Taken together they amount to nearly complete specifications with many electrical and mechanical details worked out. I am mindful of the adage that a camel is a horse designed by a committee. With some luck, however, we may be able to avoid the greater grotesqueries of bureaucratic design.

The ultimate aim of your work is a self-contained array of PbS detectors including filters, chopper, chopper driver and as much of the electronics as practicable. Eight channels is the preliminary estimate of the required number by Lane Kelley.(1) Based on the work by John Adams and his colleagues at MIT and the University of Washington, I am not sure that the number shouldn't be even higher, say 16 or so. The larger number of channels would involve a number of mechanical problems, some of which we have already discussed but all of which seem to me to be clearly out of the range of the present work. For immediate requirements, however, I think that two channels are sufficient, provided that the design is done in such a way as to allow for direct scaling to eight channels. This includes provision for such items as channel selection, circuit real estate, chopper blade size, number of feed throughs, etc. Working out the problems of two channels and implementing them is as much effort as can reasonably be expected in this year. Sufficient unto the day is the evil thereof.

There has been persistent and, to me, puzzling discussion of the channel distribution and operating modes for the array. There is no virtue and there are considerable disadvantages to trying to combine visible channels based on silicon diodes with the PbS detectors. The silicon detectors would get no benefit from the chopper and their preamplifiers would use real estate needed for the PbS electronics. There certainly would not be room for the sixteen to twenty-four channels of near UV, visible, and near IR that a next generation system would require. Their placement with the PbS channels would merely complicate the optical system. Instead I envision that the next generation camera would use two detector arrays. For simplicity and compactness the two arrays would be mounted with their optical axes at right angles to one another. (See Fig. 1). A stationary dichroic mirror or a solenoid driven totally reflecting mirror would divert the light from the one array to the other as needed.

I also believe that the next generation camera would use in-line channel multiplexing for its signal to noise advantage and will probably use an imaging rather than a so-called "spectrometric" mode of operation. In the latter mode, the camera mirror is positioned and the detectors are stepped through the channels. There are really only two justifications of such operation: reduced data rate for communication or storage purposes and increased signal to noise by using a very slow scan rate without the penalty of ultra long data acquisition times. For reasons discussed below the PbS detectors will probably be much larger than the Viking photodiodes. The consequence of this is that imaging coverage is possible in all eight channels in less time than for a single channel at the rather modest Viking data rate of 16 kbps. Going to within the line as opposed to line sequential channel multiplexing makes possible multiple channel imaging with reasonable data rates and with reduced requirements on channel settling times, while maintaining the option of using a spectrometric mode purely for its signal to noise properties. I envision then a final array having two switchable sets of four channels active at a time. There would be four amplifier/demodulator combinations with low level switching between two PbS detectors for each amplifier. For the first cut at this, one would have two complete channels with provision for testing out the low level switching for channel isolation and for settling time.

By way of concluding prefatory remarks, let us set down some of the relations between noise equivalent power, response time, and responsivity which hold for PbS detectors. The object of this is to provide a common notation, to give a simple view of how the predominant design criteria interact and to give some idea of the numerical values possible with commercial detectors. Because the principle noise limit in photoconductors is statistical fluctuation in the number of carriers in the conductor, the noise is proportional to the square root of the total number of carriers in the material. For a given type and thickness of material, it is proportional to the square root of the area. It is customary to specify a detectivity $D^*$ which is independent of detector size as

$$D^* = \frac{\sqrt{A\Delta f}}{NEP} \qquad (1)$$

where NEP is the noise equivalent power
A is the detector area
$\Delta f$ is the measurement bandwidth.

The implicit hope behind including the factor $\sqrt{\Delta f}$ in the formula for $D^*$ was to make $D^*$ independent of the frequency of chopping below the response time of the detector. Since the lead salt detectors are dominated throughout their useful response range by $1/f$ noise, it is in fact necessary to specify the chopping frequency at which the measurement was done, and customarily the chopping frequency for highest $D^*$ is chosen. Since the response of the detector is wavelength dependent, it is also necessary to specify the kind of optical power to which the electrical noise is equivalent, whether monochromatic or black body ratiation, and if monochromatic, at what wavelength. For our purposes the most useful detectivity is that for the peak monochromatic response which occurs near 2 μm.

The temporal response of a photoconductor is related to the recombination time for the photogenerated carriers. In effect the detector cannot sense disappearance of illumination any faster than the excess carriers spontaneousl- dissappear by electron-hole recombination. Thus the temporal response is given by

$$H_{PbS}(s) = \frac{1}{1+s\tau} \tag{2}$$

where $s$ is the usual Laplace variable
$H_{PbS}$ is the transfer function of the detector
$\tau$ is the recombination lifetime.

There is also a relationship between detectivity and $\tau$ which is important to the designer. Because generation recombination processes dominate the noise in room temperature detectors, it is usually true that

$$D^* \propto \tau \tag{3}$$

The circuit designer needs to know the responsivity of the detector and the noise voltage level at the detector. From first principles one can show that for a simple photoconductor at or below the peak response wavelength:

$$R = \frac{\eta\lambda\mu\rho_{\square}\,\tau}{1.24\cdot A^{1/2}}\,V_{app} \,.$$

where $R$ is the responsivity in volts/watt at wavelength $\lambda$ in microns
$\eta$ is the quantum efficiency
$\mu$ is the carrier mobility
$\rho_{\square}$ is the resistance between two opposite edges of a square detector
$V_{app}$ is the voltage applied across the detector.

The response voltage is the change in voltage across the device assuming constant current through the detector. Generally the quantum efficiency and mobility are not given separately and what is important to the designer is the functional relationship that

$$R \propto \frac{\lambda\tau\rho_{\square}\,V_{app}}{A^{1/2}} \tag{4}$$

since the constant of proportionality will often be the same over the whole range of a given manufacturer's products. Combining the respositivity with the detectivity leads to the expression for noise voltage density

$$V_N = K_N\,\frac{\tau\rho_{\square}\,V_{app}}{D^*} \tag{5}$$

where $V_N$ is the noise voltage density in volts/$Hz^{1/2}$
$K_N$ is a constant.

The usefulness of (5) comes from the implication that front end design can proceed independently of detector selection. Usually $V_{app}$ is constrained by available voltage, and heating of the detector. The usual procedure for design of the low end stages is to design for a gain that will raise the noise voltage to a level comparable to the drift which has to be anticipated in the demodulator and subsequent stages. Since $\tau/D^*$ is a constant for a wide range of detectors and since $\rho_{\square}$ is well-controlled, equation (5) allows one to design an amplifier without preselecting a $D^*$ or $\tau$.

It is not particularly easy to get the value of $K_N$ in equation (5) from manufacturer's data sheets, but I have attempted to do so for typical detectors from three manufacturers. The results are shown in a table in Fig. 2.

Let us now consider some actual design proposals.

1. Detector area: The single parameter which most effectively controls the signal to noise ratio of the array is the area of the detector. As a criterion for the selection of this parameter, I suggest that we simply design for the same sensitivity in a Viking like camera as the Viking photosensor arrays had. This criterion has the advantage of being simple to use and it yields an immediately useful device. Let the sensitivity be defined as the NEP divided by the detector area, i.e. the RMS noise equivalent power per unit area. Let the subscript V indicate Viking and the subscript PbS, the new array. Then for Viking at about .6μm with a noise bandwidth of 2.8 kHz we had NEP of $1.45\cdot10^{-12}$ watts; and the detector area was $1.23\cdot10^{-9}\text{cm}^2$. Thus

$$S_V = 1.19\cdot10^{-8}\text{watts/cm}^2$$

For the lead sulfide detectors we have

$$S_{PbS} = \frac{1}{D^*}\left[\frac{\Delta f}{A_{PbS}}\right]^{1/2}$$

The required bandwidth is reduced in proportion to the number of detectors to be multiplexed in line since one would slow the mirror down to maintain constant data rate and to take advantage of the corresponding reduced bandwidth. The bandwidth is also reduced in proportion to the detector size; in the Viking camera the required bandwidth would have been only 1.7 kHz if attention had been paid to the proper shaping of that response. Thus

$$\Delta f = \frac{1700}{N}\left(\frac{A_V}{A_{PbS}}\right)^{1/2}$$

and equating the sensitivities of the two arrays yields:

$$A_{PbS} = \left[\frac{1}{D^*S_V}\bigg/\frac{1700A_V^{1/2}}{N}\right]^{4/3}$$

where N is the number of multiplexed channels.

Using the parameters for the Infrared Industries type 2 device, with $D^* = 4.10^{10}$ and $N = 2$ yields

$$A_{PbS} = 1.24 \cdot 10^{-3} cm^2$$

or a square detector .015" on a side. The corresponding bandwidth is 283 Hz.

There are several possible objections to this procedure. There is less energy in the solar spectrum at 2μm than at .6μm by a factor of (3). There are more problems on Earth with atmospheric absorption. Also the Viking system has a marginal signal to noise ratio for radiometric measurements. However, higher signal to noise ratios require still larger detectors with concommitant problems in degraded spatial resolution and in more stringent chopper requirements.

Alternatively one might argue that $D^*$ is chosen too low since other detectors go to 1 to $1.5 \cdot 10^{11}$. Unfortunately, by equation (3), the higher $D^*$ is bought at the cost of increased response time while the implications of the reduced size would be an increased bandwidth requirement. For the detector used in the example, the cutoff frequency is about 1.3 kHz. The detector must respond to signals in a band ±283 Hz about a center chopping frequency which might then be 1 kHz. To try to improve spatial resolution with a longer response time detector quickly runs into the problem that twice the bandwidth of the required signal exceeds the bandwidth of the detector.

On the whole, I feel a detector of .015" square and a response time around 120 μsec offers the best compromise. If performance requirements dictate higher S/N then further reduction in scanning speed can be tried and in the final unit that can be combined with a higher $\tau$ detector and a slower chopper.

2. Detector spacing: Detector spacing is controlled by two considerations. First, how large is the cone of convergence of the rays from the lens. Second, how much space must be allowed between holes in the filter mask. The convergence half angle is given by $\tan^{-1} \frac{1}{2f}$ where $f$ is the aperture number (ratio of focal length to diameter) of the lens. Although cameras may vary in lens focal length, the range of $f$ number is smaller. The Viking camera used an f-5.65 lens. Further improvement in S/N and spatial acuity can be gotten by using lenses to f-4. Below that lens design becomes difficult and the optical system grows rapidly in bulk. I suggest designing for an f-4 system as worst case. Figure 9 shows the ray diagram from which the detector spacing is calculated. For the spacing between mask holes I rather arbitrarily chose .012". This is based on allowing a .004" kerf region on each side of the top filter edges and a filter thickness of .020". As discussed below under filter holder design, I suspect it may be necessary to use a silicon electrostatic shield under the filters, so I allowed a generous .030" below the filter holder as .020" for clearance above the wire bonds and .010" for the silicon. If no silicon shield is used, then the .010" can be used to stiffen the filter holder. The resulting detector separation is .020". For 2 detectors the array length is .050" long; for 8 detectors .260" long.

3. Chopper selection: Chopper frequency should be as high as is compatible with the detector response time and the required blade size. The purpose behind the high frequency is to minimize the 1/f noise and the size of the chopper. The required blade size is determined by the array size and the height between the blades and the array. I discussed the latter problem with the Bulova people and they tell me that by putting the blades at the bottom of the tines and setting

the tines low on the pole pieces that the blades could be spaced .120" above the mounting plate. Allowing .035" for the clearance between the mounting plane and the detectors leaves .155" from the blades to the detector. The required maximum optimum opening is the width of the detector plus slightly less than the height divided by f. (Slightly less to maximize the fundamental component in the chopped signal.) I get a number near .048". Naturally, at rest, the blades would be half that. In length, the blades need to be slightly longer than the array length plus the height divided by f to allow for placement tolerance. I make it about .320". These dimensions are well within Bulova's capabilities at frequencies to 1.3 kHz. The blade finish should be bright, preferably gold, to reduce emission as the chopper warms up.

As remarked above, with a 120 μsec detector, one could use a 1 kHz chopper and still have the 3 DB response point above the upper sideband. However, it might be wiser to go slightly lower in frequency, to say 800 Hz. At that frequency the 1/f noise is only slightly worse (factor of 1.1) but the 3 DB response is above the upper sideband in the worst case for that detector series. Also, by slightly fooling around with the amplifier response shape, one might be able to use the next higher detectivity device, which has a 200 μsec response time.

4. Front end amplifier: Two decisions are critical in the choice of front end design. They are the selection of circuit configuration and the selection of an input amplifier device. The traditional configuration is a non-inverting amplifier capacitively coupled to the PbS and a biasing resistor. This circuit is probably not very suitable for us for a number of reasons: it does not lend itself well to low level multiplexing of several channels; it requires higher bias voltages for a given signal level; it allows only very crude shaping of the frequency responses. Instead I recommend the somewhat novel circuit of Fig. 3a. This uses a non-inverting stage to keep the input node voltage very low and an integrator around the input stage to shape the frequency response and to maintain the bias within the proper range. The transfer function assuming ideal amplifiers would be:

$$H(S) = -\left(\frac{R_1}{R_D}\right)\left[\frac{S\tau_2'}{S^2\tau_1\tau_2'+S\tau_2'+1}\right]$$

where $\tau_1 = R_1C_1$

$$\tau_2' = R_2C_2\cdot\left(\frac{R_3}{R_1}\right)$$

This is a bandpass amplifier having a center frequency of $f_c = \dfrac{1}{2\pi\sqrt{\tau_1\tau_2'}}$, a gain of $R_1/R_D$ at that frequency and a Q of $\sqrt{\tau_1/\tau_2'}$. The advantages of this circuit are the following:

1) It allows low level multiplexing of an indefinite number of channels by switching bias voltage. Since the input node is always near zero potential regardless of detector resistance or illumination, there will always be good channel isolation no matter how many channels share the switched bias line.

2) It has low sensitivity to input voltages from chopper capacitance because of the low node voltage.

3) The response function can be used to shape the overall band shape in a desirable way.

4) It has low sensitivity to noise on the negative power supply lead.

The disadvantages are:

1) It uses an extra amplifier. This is not serious since the remaining operations need more amplifiers and everything can be accommodated with one quad operational amplifier chip per channel.

2) The circuit phase margin is very tight. Further checks by computer modeling on the effects of high frequency amplifier poles on circuit stability are necessary.

We defer to a later section the selection of gain, center frequency and Q.

The selection of the A1 device type is purely a matter of optimum noise performance. Figure 3b shows a noise model for the stage from which one can see by inspection that the voltage noise density (per root Hz) at the output of the amplifier is

(6)

$$V_{\text{noise out}} = \left[\left(R_1 i_n\right)^2 + e_{OA}^2 R_1\left(R_D+R_3\right)/R_D R_3 + 4kTR_1\left(1+\left(\frac{R_1}{R_D}\right) + \left(\frac{R_1}{R_3}\right)\right) + V_{nD}^2\right]^{1/2}$$

where $i_n$ is the amplifier input noise current; for an amplifier with no internal bias and negligible 1/f noise $i_n^2 = 2ei_{bias}$

$e_{op\ amp}$ is the noise voltage density of the operational amplifier referred input

$V_{nD}$ is the detector noise given by equation (5) above.

The choice of amplifier type reduces to minimizing the sum of the first two terms in the brackets. Figure 4 gives a table of these terms for several candidate devices with typical resistence choices. The table shows an even tradeoff between the LM108 and the LM216A, other types being clearly inferior. On the whole, the LM108 is the better choice. The bias voltage $V_{app}$ is chosen to make $V_{nD}^2$ the dominant term in (6) subject to the constraint that it be low enough to be switchable and not to heat the device. The amplifier supply voltage between 12 and 15 volts would be suitable. Then $R_3 = \frac{1}{2} R_{D\ nominal}$ would insure the proper bias range for the operation of A2.

5. <u>Second stage and demodulator configuration:</u> Some extra gain before the demodulator is probably desirable, and making that gain a tuned stage substantially improves the possibility of shaping the overall frequency response. The demodulator has to be a full wave type to maximize the signal to noise and to make the ripple components as small as possible. The output has to be driven from a low impedance source. Probably to reduce the ripple components sufficiently one would use a low pass filter of third order or higher, but it is not necessary or desirable to include more than the first pole of that filter within the array. A simple circuit which meets these criteria is shown in Fig. 5. It is by no means the only possibility, but I think it would probably work well.

Amplifiers A2, A3, A4 and A5 can all be part of an LM124 chip. The FET switches might be a Siliconix DG200 chip but I expect you have a preferred device type from your other work. The transfer function of the tuned stage is given by

$$H(S) = -\frac{R_6}{R_4}\,\frac{S\tau_4\left(1+S\tau_5\right)}{\left(1+S\tau_4\right)\ \left(S^2\tau_5\tau_6+S\tau_5+1\right)}$$

where $\tau_4 = R_4C_4$

$\tau_5 = 2C_5R_5$

$\tau_6 = \frac{1}{2}\,C_5R_6$

Adjusting $T_4 = \tau_5$ for pole cancellation leads to a transfer function for a bandpass filter with gain $= \left(\frac{R_6}{R_4}\right)$; center frequency $= \frac{1}{2\pi C_5\sqrt{R_6R_5}}$; and $Q = \sqrt{\frac{R_6}{4R_5}}$.

6. <u>Component selection:</u> In considering the filter design, I realized that it made sense to use the bandwidth for the final 8 channel array so that the design and testing of the filters would not have to be repeated. The argument for higher signal to noise seemed compelling enough that I did not recalculate the detector area but simply assumed that the factor of 2 in bandwidth will result in a 40 percent improvement in noise. I assumed further that the MTF of the system would be dominated by the $\sin\omega/\omega$ cutoff due to the detector size and that the velocity of the mirror would be chosen to place the first zero of this at four times the bandwidth out from the center frequency. (This corresponds to pixel spacing equal to the length of the detector element and a bandwidth equal to half the sampling rate.) I adjusted a second order transfer function to optimize the MTF. Fig. 6 shows the resulting low pass equivalent transfer function. Applying a bandpass transformation yields Fig. 7. The corresponding requirements for center frequencies are 687.24 Hz for the first stage and 931.26 Hz for the second. The Q of both stages is 5.76. I chose a first stage gain of 15 somewhat arbitrarily and a second stage gain of 10 to get the estimated RMS noise up to about 3 mV as being safely above drift levels in subsequent stages. The design equations for the amplifiers (<u>vide supra</u>) then lead to the circuit drawn in Fig. 8. The detector is assumed to be the Infrared Industries Type 2 series.

If a longer response time detector is used then it is possible to choose unequal Q's for the two stages and remove the response time effects from the overall MTF curve. Naturally the circuit has to be computer modeled for stability and for the correctness of component values before it is implemented.

7. Miscellaneous remarks on the case: The case should be made from the easiest, cheapest material to work with. I recommend brass as being easily machined and easily soldered for feedthrough, yet durable and tough enough for the job. Nickel plating with or without gold would be nice. We do not have any environmental constraints to make Kovar necessary. Likewise, there is nothing sacred about a round case - rectangular will do. Similarly, it may be easier to use a square window. I recommend Schott RG780 glass for the window on the grounds that it will do a good job of blocking visible light from the array.(4) To seal the array, I think a clamp or some screws can hold it together, but probably a little of the 2-part silicone RTV sealant on the outer edge seal might be prudent for final shipment and/or test.

It seems to me that having filters made for the array on .010" to .020" glass is likely to be expensive and to be a major headache. I suggest holding off getting that done until the main job of the electronics and hybrid packaging has been planned and started. If funds then seem available I suggest asking Coherent Radiation and OCLI for a best effort, single shot try at a .1 - .12 μm filter centered at 2 μm on a .020" glass of their choice. This would be minimum cost and would get us something to use and something to practice cutting on. In the matter of cutting I suspect that one could get better results using photoresist to mask an etch of kerf paths through the interference layers before cutting. Done right, this might even allow one to scribe and break the substrate instead of having to cut.

I think the filter holder can be flat and should be quite large, extending partway under the chopper and out completely covering the electronics. Such an extension simplifies the reduction of coupling from the chopper into the preamplifiers and lowers the amount of scattered light inside the array. A lip or rim around the inside wall of the case can prevent the mask from falling into the electronics during assembly. The material might be .010" stainless steel or brass shim stock overcoated with Ni/Au/black chrome. Electron discharge machining is probably necessary to make the precision holes for the filter aperture. If LRC cannot find a supplier, I may be able to find one here.

A final problem with the filter holder is the possibility that the chopper may couple electrostatically through the filters to the detectors. The traditional solution is to interpose a conducting, grounded, transparent surface between the two. The traditional tin and indium oxide coatings are opaque in the infrared so some other material is necessary. One possibility is for me to make you some thin bars of 10-ohm cm silicon with aluminum contact rims. I would give both sides an antireflection coating of quarter wave $SiO_2$ peaked at 1.8 μm, and you could mount them to the underside of the filter holder with conducting epoxy, preferably silver epoxy. If you want some of these, even just as a precaution, let me know the exact dimensions. To save space, one would probably recess the silicon into the filter holder.

The following final remarks are in no particularly logical order. Finding wire bondable feedthroughs may be a problem, for which there are a couple of solutions. One possibility is to use IC headers soldered into the case. I could get some of those. Another possibility is to use solder-plated feedthroughs and use indium solder on gold wires bonded to the mother board. Tricky, but possible, and certainly better than trying to plate up your own feedthroughs as EG&G did.

Separate power lines including ground are probably needed for the chopper drive circuit to avoid coupling problems.

Lead sulfide detectors do not like ultra violet light and take several days to recover, even from extended exposure to fluorescent light. They are also intolerant of cleaning chemicals and high temperatures such as for epoxy curing.

8. Summary: This note contains enough material that it may be difficult to digest. I would enjoy talking it over with you and hope that such a conversation might dispel any suspicion of hubris in me. To pull the material together, the following is a table of the principle array characteristics:

| | |
|---|---|
| Type of detector: | room temperature lead sulfide; Infrared Industries Inc. type 2 or type 3. |
| Number of channels: | two channels active at all times with separate bias connection for each one. |
| Size of detectors: | .015" square on .035" centers. |
| Bandwidth: | 140 Hz. |
| Preamplifier circuit: | see Fig. 8. |
| Chopper: | Bulova type L2-C with frequency 800 Hz and bright finish blades mounted as closely as possible to the base plane. The blades are .310" long with a rest separation of .025" nominal. |
| Case material: | nickel plated brass. |
| Window: | Schott RG-780 glass. |
| Case seal: | silicone rubber if needed. |

Yours truly,

William R. Patterson III
Senior Research Engineer

WRPIII/rw
xc: Lane Kelley
Fred Huck

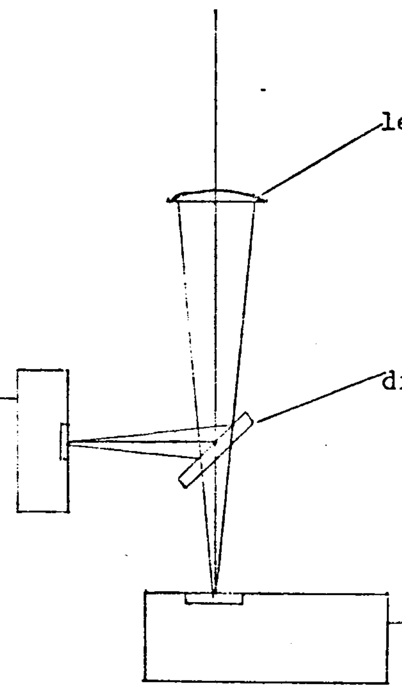



FIG. 1

| MFR. | TYPE | $\tau$ | D* | $K_N$ † |
|---|---|---|---|---|
| Infrared Industries Inc. Waltham, Mass. | type 2 | 120 μsec | $4.10^{10}\ \frac{cmHz^{1/2}}{W}$ | 20(?) |
| Optoelectronics Inc. Petaluma, Cal. | OE-20 | 350 | $10^{11}$ | 14 |
| SBRC Inc. Santa Barbara, Cal. | R. T. PbS | 400 | $8.10^{10}$ | 22 |

† see equation (5): units are cm/(watt·sec·ohm)

FIG. 2

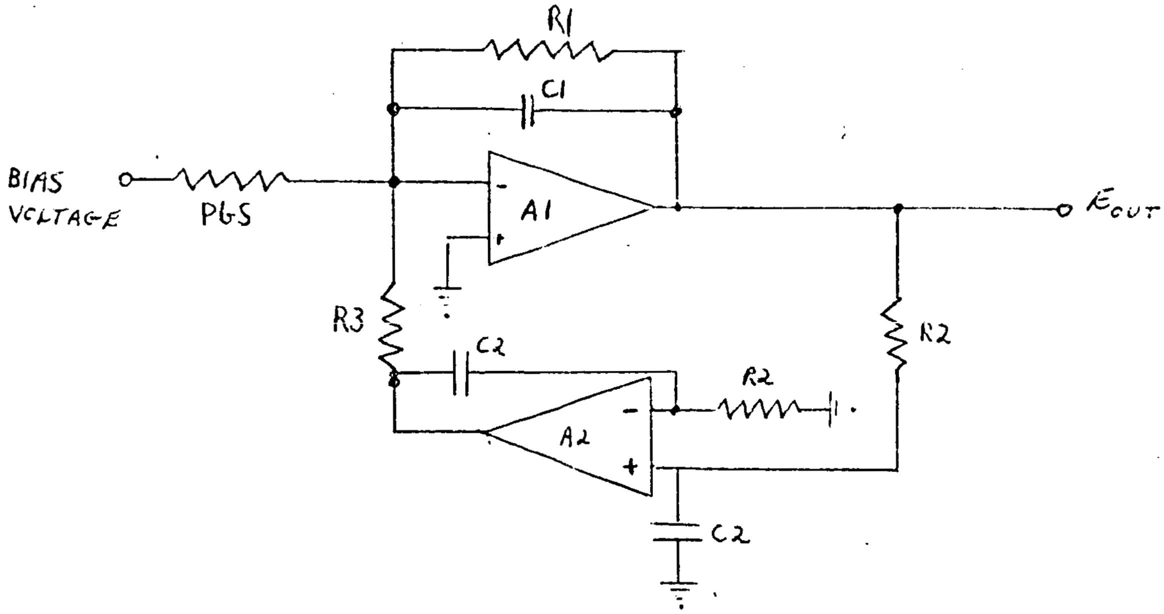
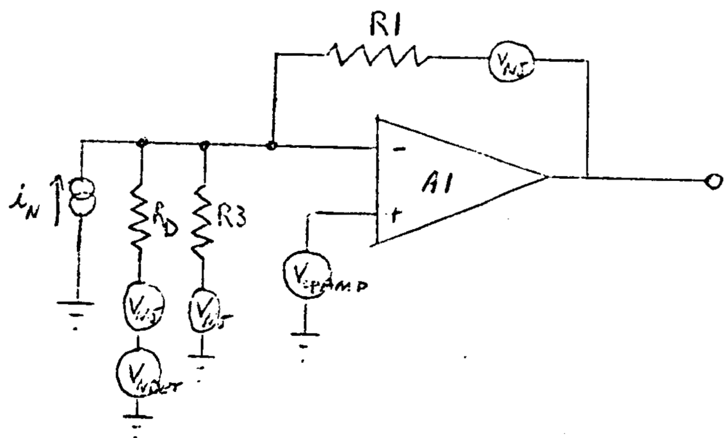



FIG. 3a




$V_{N5}$ - Johnson noise density
$V_{NDET}$ - Detector noise density

FIG. 3b

| MFR. | TYPE | $i_N R_1$ | $e_N\left(\frac{R_1}{R_3} + 1\right)$ | RSS | $I_{supply}$ | MAX. $V_{offset}$ |
|---|---|---|---|---|---|---|
| National | LM107 | 4.5 μV | 4.8 μV | 6.6 μV | 1.2 mA | 2 mV |
| National | LM108 | 2.7 | 5.4 | 6.0 | .3 | 2 |
| National | LM216A | 0 | 7.5 | 7.5 | 1.6 | 3 |
| Intersil | 8007 | 0 | 20 | 20 | 3.4 | 30 |

FIG. 4

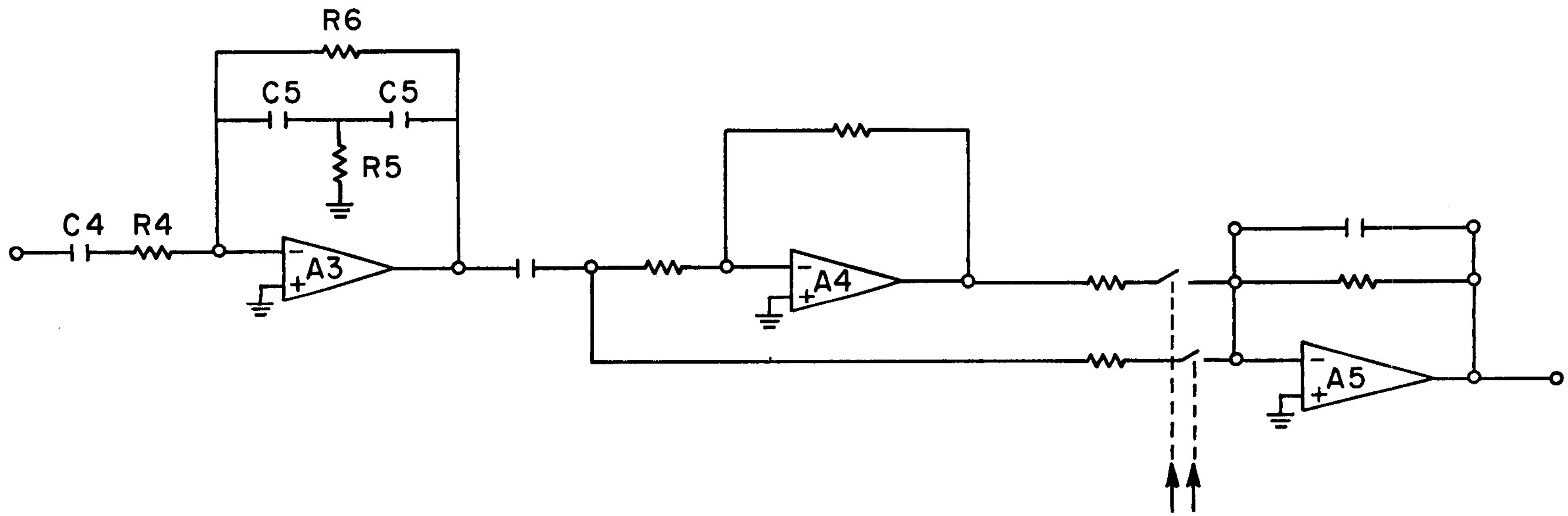



FIGURE 5

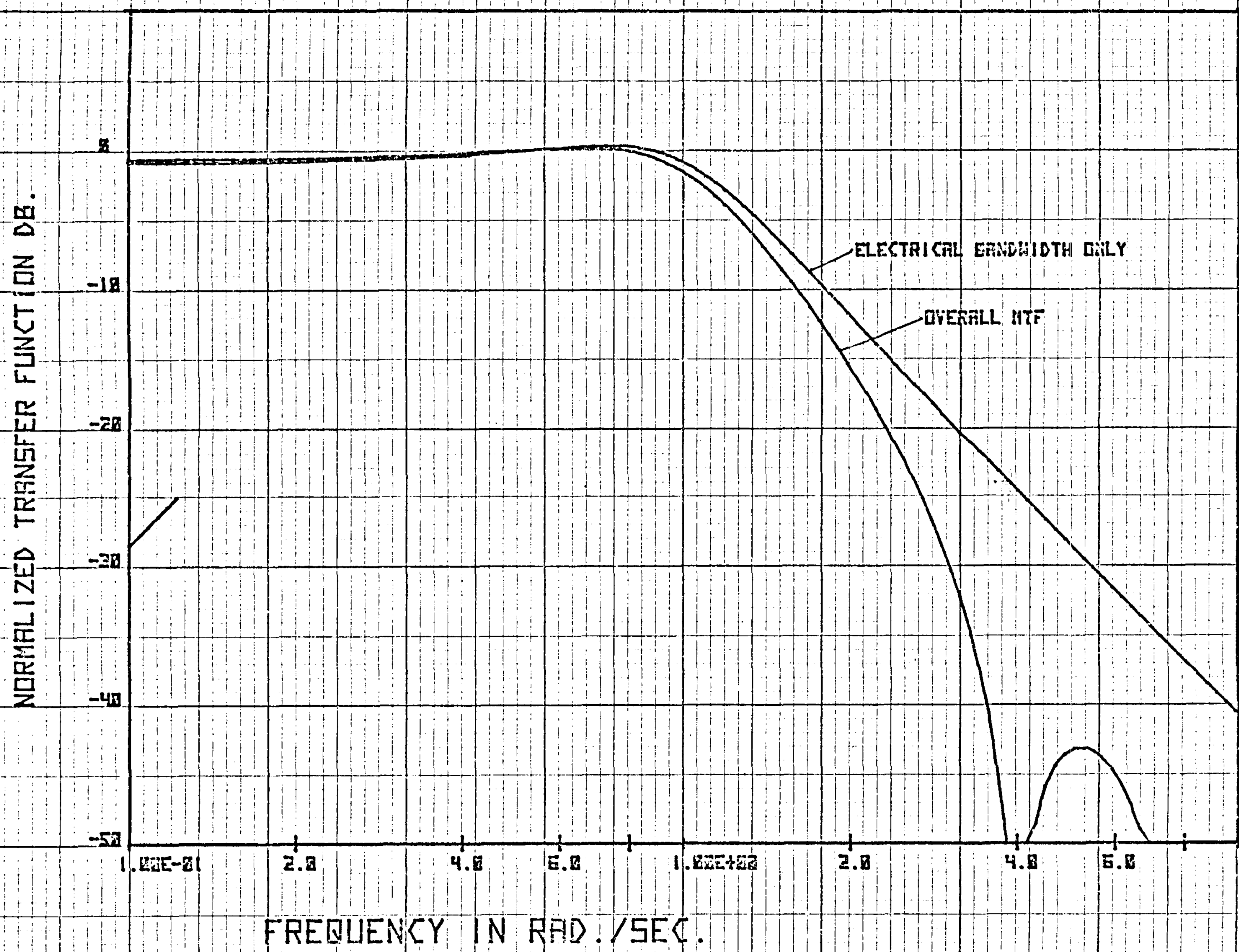

ELECTRICAL BANDWIDTH ONLY
OVERALL MTF
NORMALIZED TRANSFER FUNCTION DB.
0
-10
-20
-30
-40
-50
1.00E-01
2.0
4.0
6.0
1.00E+00
2.0
4.0
6.0
FREQUENCY IN RAD./SEC.

FIG. 6

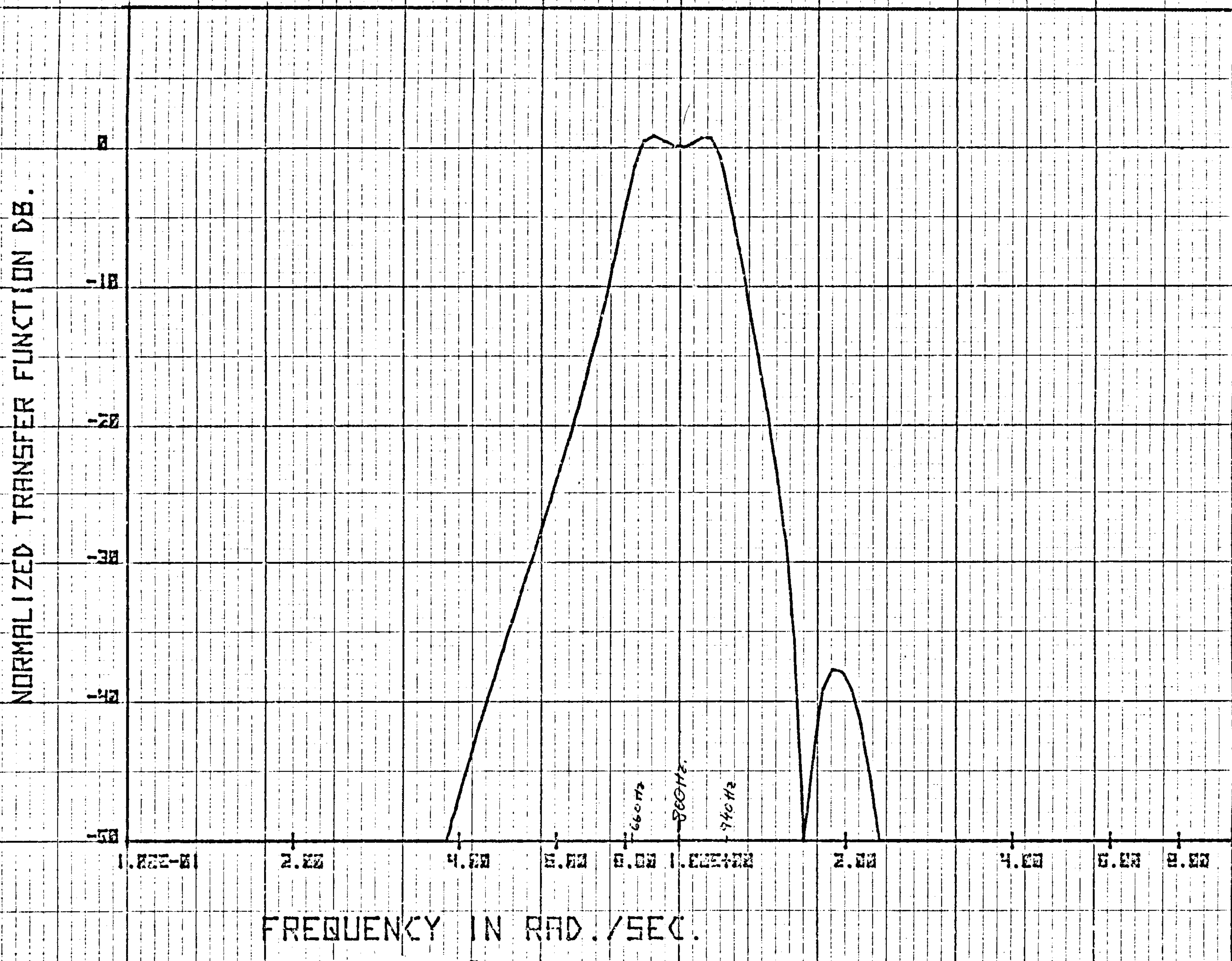
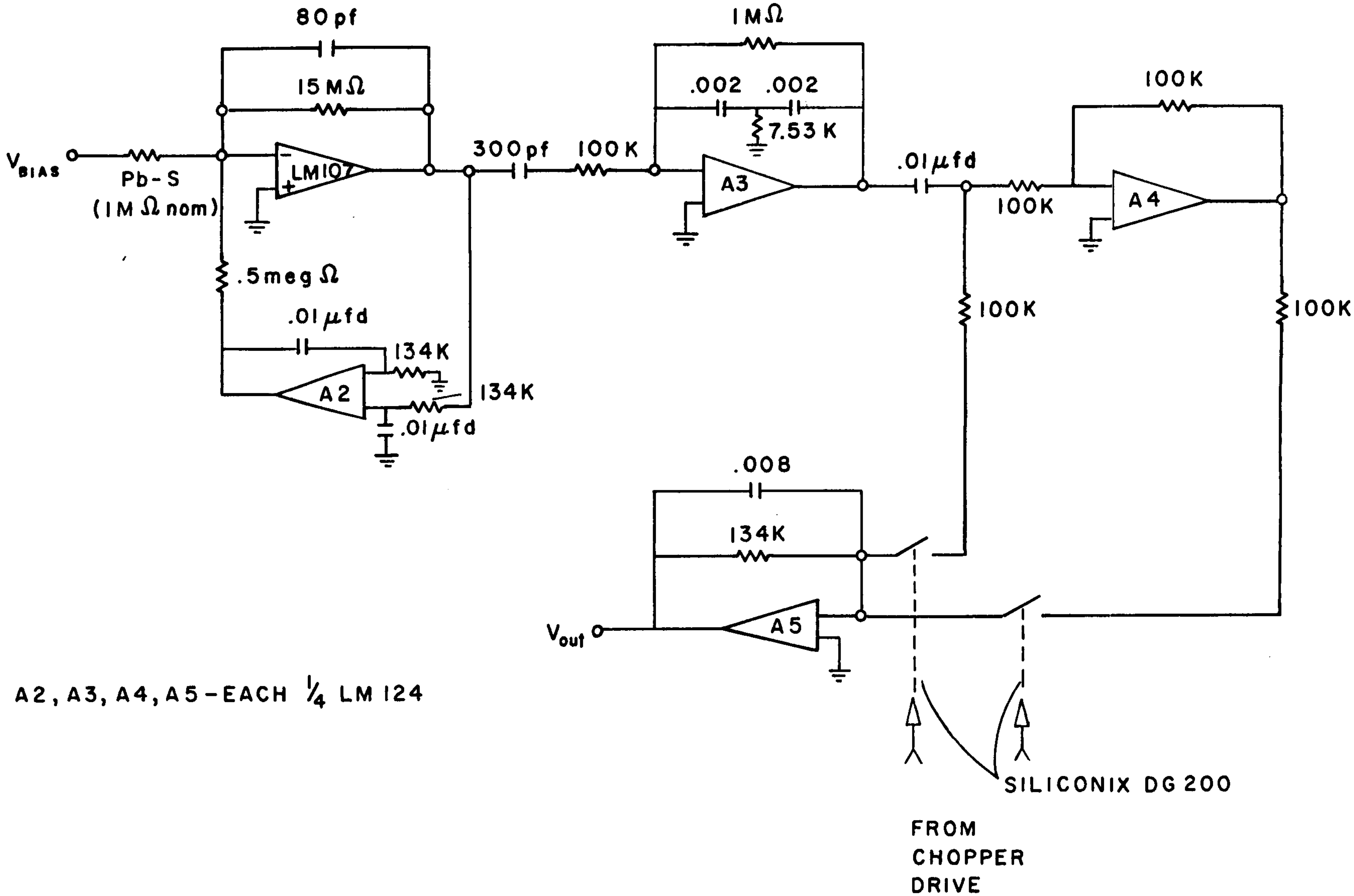
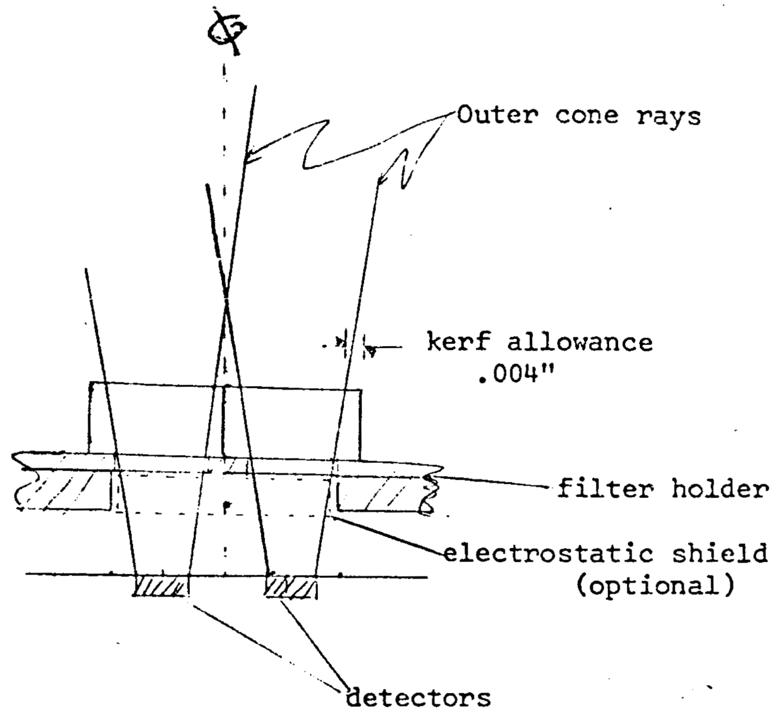

NORMALIZED TRANSFER FUNCTION DB.
0
-10
-20
-30
-40
-50
660Hz
860Hz.
940Hz
1.00E-01
2.00
4.00
6.00
8.00
1.00E+00
2.00
4.00
6.00
8.00
FREQUENCY IN RAD./SEC.

FIG. 7




A2, A3, A4, A5-EACH 1/4 LM 124

FIGURE 8



Outer cone rays
kerf allowance
.004"
filter holder
electrostatic shield
(optional)
detectors


FIG. 9

Analysis of Signal to Noise in PbS System

We wish to calculate the advantage or disadvantage with respect to signal to noise ratio that a high frequency chopper confers relative to a flag and sample subtraction technique for stabilizing a DC PbS system. We shall assume that each system has a bandwidth of 140 Hz, that the detector in each is optimized for the particular type of system, and that the detectors have sufficient bias to make the detector noise the dominant system limitation. Invariably that noise will be 'flicker' noise for which the mean square noise density has the form $K^2/f$ where $K$ is a constant and $f$ is the frequency. If $H(f)$ is the relative response of the amplifier, then the noise voltage is

$$V_N = K\sqrt{\int_0^\infty |H(f)|^2/f\,df} \tag{1}$$

To apply equation (1) to the DC system, we assume that the amplifier cuts off abruptly at 140 Hz and below that frequency has a flat response down to a frequency at which the sample and subtract system has an effect on the relative response. This frequency is approximately $1/2\,T_s$ where $T_s$ is the time between samples. We assume that this lower frequency also represents an abrupt cut off. Then

$$V_N = K_{DC}\sqrt{\ell n(f_{upper}/f_{lower})} = K_{DC}\sqrt{\ell n(280T_s)} \quad .$$

For the chopped system, we again assume that the gain is flat within the pass-band of the system. We also compensate for this approximation by introducing a form factor $B$, the ratio between the apparent bandwidth and the noise equivalent bandwidth. Thus

$$V_N = K_{AC}\sqrt{\ell n\left[B\,\frac{(f_c + 70)}{(f_c - 70)}\right]}$$

where $f_c$ is the chopper center frequency.

For a typical 2 pole amplifier after the synchromous rectifier, $B$ would be about 1.2. With a chopper frequency of 800 Hz we have

$$V_N = K_{AC}\sqrt{\ell n(1.43)}$$

If the two amplifiers have the same gain then the ratio of $K_{DC}/K_{AC}$ is the inverse ratio of the peak detectivities of the two types of detectors. It is possible to increase the response time of the DC system detector and this allows for some increase in detectivity. Comparing the use of an IR Industries type 3 with the Opto type OE-20 gives $K_{DC}/K_{AC} = .8$. Under this condition, the ratio of responsivities is simply given by the ratio of the mean absolute value of the fundamental frequency of the chopped light to the incident light level. This ratio is not especially sensitive to the chopper design, so long as the design is nearly a 50% duty cycle. For a square wave chopper it is $4/\pi^2$; for a sinusoidal chopper it is $1/\pi$. The tuning fork chopper should be somewhere in between or about 1/3. Suppose we let $Q$ be a figure of merit defined as the ratio of the signal to noise ratio of the AC system to the signal to noise ratio of the DC system. Then

$$Q = \frac{1}{3}\left(\frac{K_{DC}}{K_{AC}}\right)\sqrt{\frac{\ln(280T_s)}{\ln(1.43)}}$$

$$= .27\sqrt{\frac{\ln(280T_s)}{\ln(1.43)}}$$

For line sampling each line, $T_s$ would be ≃ 2 sec. and $Q$ would be 1.21. We have neglected the effect of the balancing PbS cell on the noise; this would introduce a factor of $\sqrt{2}$ onto $Q$ and it is probable that a $T_s$ of 12 sec, as in the Viking camera, would represent a better design because of longer mechanical life. These two factors would lead to $Q = 1.94$.

rw

Appendix II

Noise Statistics in the Viking Cameras

Consider a camera which scans a uniformly illuminated, uniformly reflecting area subtending $M$ pixels. In the Viking cameras the noise in horizontally adjacent picture elements makes the measured pixel values statistically independent. Moreover, because the noise bandwidth prior to the A/D converter is much more than half the sample limit (5 kHz vs. 1.6 kHz) and because the A/D converter is of the resetable integrator type with an integration time of only 1/3 of a pixel, I think one can treat even inline adjacent samples as roughly independent. Suppose one averages the $M$ pixels. The question then arises: how does one estimate the probable range of error in the average. The statistical properties of the average are a function of the overlying gaussian noise. We begin by deriving the probability distribution for the individual pixel values assuming a perfect A/D converter. Let all signals be scaled in units of quantization levels, i.e. if $DN = 64$ is 1 volt then all measurements are expressed in 64ths of a volt and 1 volt $\rightarrow$ 64. Let the input to the A/D converter be characterized by gaussian additive noise with standard deviation $= \sigma$. Let us subtract from the input signal a constant level of an integer number of units so as to bring the true mean of the signal within the range $\pm 1/2$. Then the probability that the input voltage $x$ lies between $x$ and $x + dx$ is $P_A(x)dx$ where $P_A(x) = 1/\sigma\sqrt{2\pi}\ \exp[-(x-\alpha)^2/2\sigma^2]$, where $\alpha$ is the fractional part of the true input signal. For an ideal A/D converter, the output is $n$ if the input lies between $n - 1/2$ and $n + 1/2$. Let $P_D(n)$ be the probability distribution of the pixel DN values. Then

$$P_D(n,\alpha,\sigma) = \int_{n-1/2}^{n+1/2} P_A(x)dx$$

$$P_D(n,\alpha,\sigma) = \frac{1}{\sigma\sqrt{2\pi}} \int_{n-1/2}^{n+1/2} \exp[-(x-\alpha)^2/2\sigma^2]dx$$

The function erf(z) is defined as $\mathrm{erf}(z) = \sqrt{2/\pi}\int_0^z \exp[-z^2/2]dz$ and can be readily computed by several techniques (see e.g., Abramowitz and Stegun, *Handbook of Mathematical Functions*, equations 26.2.16 through 26.2.19). Let $u = x-\alpha/\sigma$. Then

$$P_D(n,\alpha,\sigma) = \frac{1}{\sqrt{2\pi}} \int_{(n-\frac{1}{2}-\alpha)/\sigma}^{(n+\frac{1}{2}-\alpha)/\sigma} \exp[-u^2/2]du$$

$$= 2\left\{\mathrm{erf}\left((n+\tfrac{1}{2}-\alpha)/\sigma\right) - \mathrm{erf}\left((n-\tfrac{1}{2}-\alpha)/\sigma\right)\right\}$$

The range of $n$ is the set of integers from $-\infty$ to $\infty$; for $\sigma >> 1$ the result approaches $P_A(n)$; the function is symmetric in $\alpha$ and so the range of variables of practical interest is $\sigma\varepsilon(.05,5)$ $\alpha\varepsilon[0,1/2]$. I attach tables of $P_D(n,\alpha,\sigma)$ and plots of the mean and standard deviation of the distributions. The mean and standard deviation are calculated by

$$\overline{M}(\alpha,\sigma) = \sum_{n=-\infty}^{\infty} nP_D(n,\alpha,\sigma)$$

for the mean and

$$\sigma_{\frac{1}{2}M}(\alpha,\sigma) = [\sum_{n=-\infty}^{\infty} u^2P_D(n,\alpha,\sigma) - \overline{M}^2(\alpha,\sigma)]^{1/2}$$

for the standard deviation. The difference between $\overline{M}(\alpha,\sigma)$ and $\alpha$

represents the ultimate limit to the resolution of the mean radiance set by the quantization. In the absence of noise this limit is one half a quantizing step. For $\sigma = .05$ it drops to $\approx .4$ of a step and drops with increasing $\sigma$ until at $\sigma = .5$ the limit is .0024 of a step. Thus we are led to the conclusion that for a given level of quantization, a system with noise at the input is capable of higher ultimate resolution than one without noise. The catch, of course, is in two problems: the A/D converter is not likely to be near enough to ideal to give a linear transfer function at high resolution. Also, as noise increases, the number of samples required to achieve a given accuracy rises rapidly.

The problem of the accuracy of the A/D converter is one which cannot be estimated theoretically since it involves a question of fact, to wit the construction of a particular converter. However, two general observations are possible. First, the accuracy of the result will be best if only a few bits of low significance change in the noise range than if high order bits change. This is because almost all A/D converters use some form of resistor network to generate a voltage which is compared to the input voltage. The principle linearity problem arises from the difficulty in making the large steps in this resistor ladder precisely equal to the correct multiple of the smallest step. If no change in the high steps is involved in a conversion, then the linearity over a fractional step will be improved. For example in a 6 bit system, averaging will be more accurate for a signal with a range of DN from 26 to 30 than for a signal from 30 to 34 since in the former only bits 1, 2 and 3 change while all six bits change in the latter. Second, changes in a signal level are more accurately determined by averaging than is the absolute value.

The question of how many pixels must be averaged to achieve a given level of accuracy is amenable to mathematical treatment. Suppose we define the error in a measured average of $M$ pixels, $\varepsilon_M(M,\alpha,\sigma,x)$, by saying that it is the sum of two terms, namely a statistical uncertainty in the estimate of the average of the DN's and the deviation of the DN average from the true average as computed above. Although the error depends on $\alpha$ in a complex way, we do not know $\alpha$ *a priori* and, therefore, we are interested in the maximum value of $\varepsilon_M$ over the range of $\alpha$ from 0 to 1/2. That is, we write

$$\varepsilon_M(M,\sigma) = \underset{\alpha\varepsilon(0.5)}{\text{Max}} \left[f_M(\alpha,\sigma,x)\sqrt{\xi_M(\alpha,\sigma)} + \alpha - \overline{M}(\alpha,\sigma)\right]$$

where $f_M(\alpha,\sigma,x)$ is a function depending on the particular distribution of DN's. The term $f_M\sqrt{\xi_M}$ represents the range in uncertainty of the DN average. There is a probability x, near unity, that the true DN average lies between the measured value minus $f_M\sqrt{\xi_M}$ and the measured value plus $f_M\sqrt{\xi_M}$.

$\xi_M(\alpha,\sigma)$ is the variance of the sum of $M$ pixels.

It is straightforward to show that $\xi_M(\alpha,\sigma) = \frac{1}{M}\xi(\alpha,\sigma)$.

The function $f_M$ can only be determined exactly from the full probability distribution function for the sum of $M$ pixels. If $P_M(M,\alpha,\sigma)$ is the probability that the sum of $M$ pixels lies between $M \pm \frac{1}{2}$ then (see Lee, *Statistical Theory of Communications*, p. )

$$P_M(n,\alpha,\sigma) = \sum_{a_1=-\infty}^{\infty} \sum_{a_2=-\infty}^{\infty} \cdots \sum_{a_M=-\infty}^{\infty} P_D(n-a_1-a_2 \cdots a_M,\alpha,\sigma) \prod_{i=1}^{M} P(a_i,\alpha,\sigma)$$

Since this distribution is impractical to calculate for all but the smallest ensembles of pixels, we must approximate it. Two limiting conditions are apparent: For $\sigma = 0$ the form of $f_M$ is unimportant since $\xi(\alpha,0) = 0$ and $\varepsilon_M = .5$. For $\sigma \geq .5$, the gaussian noise dominates the system, making $f_M(\alpha,\sigma,x) \approx \text{erf}^{-1}(x)$ and $\alpha - M(\alpha,\sigma) \approx 0$. Then we write

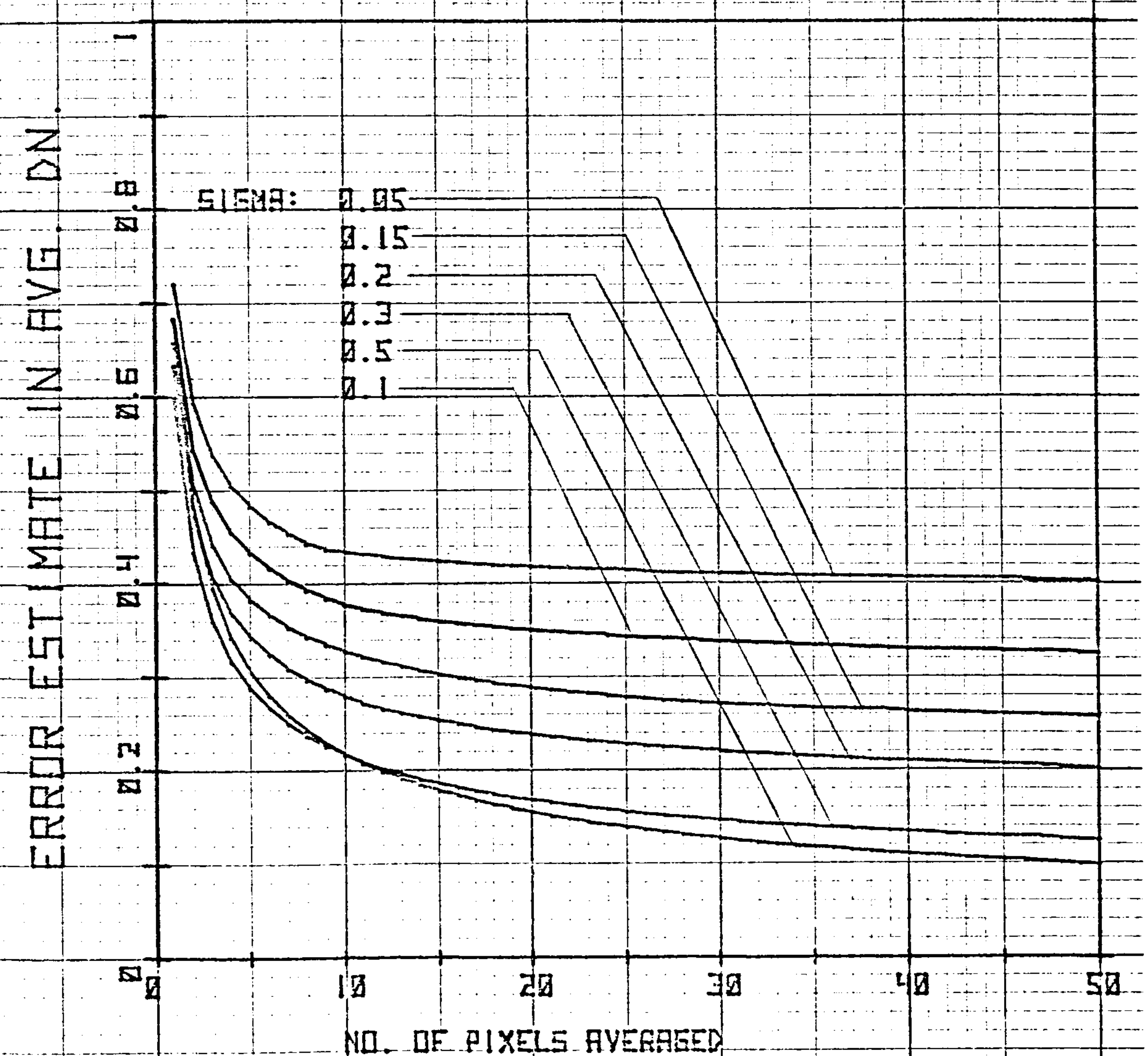
$$\varepsilon_M(\sigma,x) \approx \underset{\alpha\varepsilon(0,.5)}{\text{Max}} \left(\text{erf}^{-1}(x)\sqrt{\xi(\alpha,\sigma)/M} + \alpha - \overline{M}(\alpha,\sigma)\right)$$

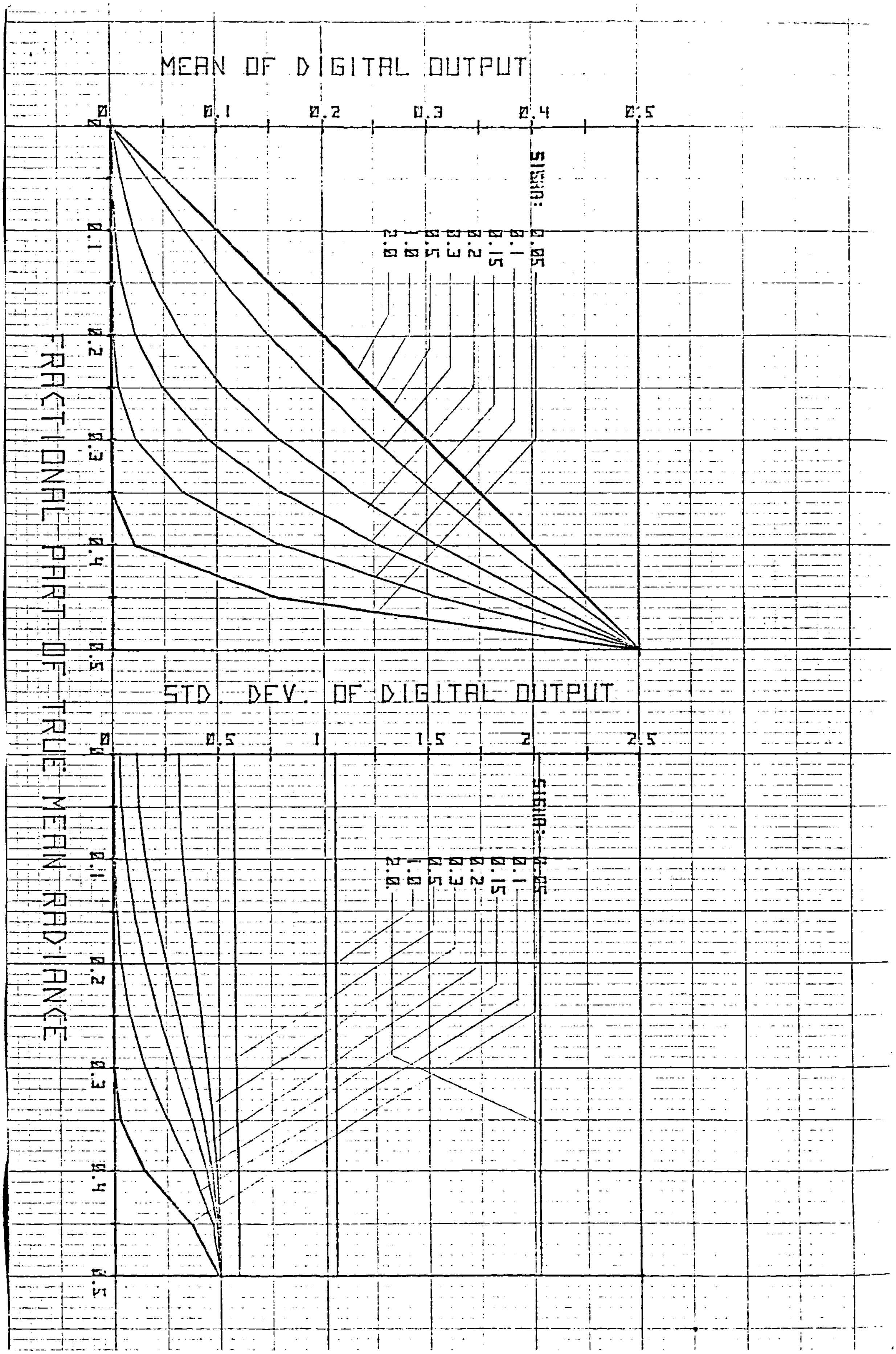
For $\sigma \geq .5$
$$\varepsilon_M \approx \text{erf}^{-1}(x)\,\frac{\sigma}{\sqrt{M}}$$

For the range of $\sigma < .5$ we have graphed $\varepsilon_M$ as a function of $M$.

These results have a bearing on the strategy employed to optimize the radiometric data. Any such attempt must reconcile the need for dynamic range (i.e., low gain) with the need for resolution (i.e., high gain). In the case where maximum resolution is required, our results suggest that one should increase the gain until $\sigma \approx .5$, which occurs between gains 1 and 2. For an ideal A/D converter, no further improvement results from a further increase of gain. In practice then the optimum gain would be determined by the number of pixels to be averaged. Only when the number of pixels to be averaged is large enough to reduce $\varepsilon_M$ to below what can reasonably be gotten from the A/D converter is an increase in gain profitable. For the gain 1 to gain 0 changeover, this point is somewhere between 16 and 100 pixels. Should it be necessary to determine the optimum more accurately, an experiment should be run on the STL lander.

WRP/jb


ERROR ESTIMATE IN AVG. DN.
SIGMA: 0.05
0.15
0.2
0.3
0.5
0.1
0
0.2
0.4
0.6
0.8
1
0
10
20
30
40
50
NO. OF PIXELS AVERAGED


MEAN OF DIGITAL OUTPUT
Ø
Ø.1
Ø.2
Ø.3
Ø.4
Ø.5
SIGMA: Ø.Ø5
Ø.1
Ø.15
Ø.2
Ø.3
Ø.5
1.Ø
2.Ø
STD. DEV. OF DIGITAL OUTPUT
Ø
Ø.5
1
1.5
2
2.5
FRACTIONAL PART OF TRUE MEAN RADIANCE